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High-frequency, high-power transistor linear power amplifier design.

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HIGH-FREQUENCY, HIGH-POWER TRANSISTOR LINEAR POWER AMPLIFIER DESIGN

by

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UNITED STATES
NAVAL POSTGRADUATE SCHOOL

THESIS

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LINEAR POWER AMPLIFIER DESIGN

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[redacted]

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ABSTRACT

This thesis discusses in detail the physical and electrical characteristics of high-frequency, high-power transistors, and why class B amplifiers are necessary for linear power amplification of signals containing more than one frequency.

Linear power amplifier design is contingent upon having a suitable design technique. "Suitable" often means being able to determine parameter values called for by that technique. Conjugate impedance matching is a suitable technique and three of the four parameter values can be accurately determined. In some cases manufacturers provide data for this technique, titled "Large-signal parame-

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I. INTRODUCTION

Single-sideband, high-frequency radio communications offer significant advantages to the military establishment. The major advantages are highly efficient use of transmitter power and conservation of the frequency spectrum. Single-sideband transmitters and receivers are considerably more complex electronically, than the conventional AM transmitters they replaced. Present techniques for producing single-sideband signals are low power level techniques. Subsequent amplification of the single-sideband signals to the desired transmitter power level must be accomplished by linear amplifiers.

Transistors have always had limited high-frequency, high-power capabilities. Much developmental effort has been expended to increase the power and frequency limitations, and considerable progress has been made (see Figure 1 and Table 1). A transistorized high-frequency linear power amplifier may offer advantages such as small size, ruggedness, light weight, and lack of standby power requirement over a vacuum-tube model. Nearly all of the limiting factors in a transistor power amplifier are the limiting factors of the transistors themselves. To achieve the desired results normally requires full utilization of the transistors' capabilities. In practical terms this means the high-frequency, high-power transistors are operating at full capability. This in turn means that the transistors' capabilities and limitations must be well known to the de-

sign engineer.

The design techniques for "normal" transistor design do not carry over well into high-frequency, high-power, linear power-amplifier design, though they are useful for visualization purposes. A technique for which parameter values may be determined is proposed after a lengthy discussion of the device with which we are concerned, namely the high-frequency, high-power transistor, and of the manner in which it is to be used, namely class B operation.

Many of the statements in this thesis are very closely related to other statements, differing primarily only in viewpoint, so that it is difficult to include the proper amount of cross referencing. The amount of cross referencing that for one person ties the differing viewpoints together may for another person be needlessly distracting. This becomes a crucial problem in the design discussion; essentially the problem is how simple does the discussion become, when the simplicity is gained by not mentioning all of the conditions and qualifications.

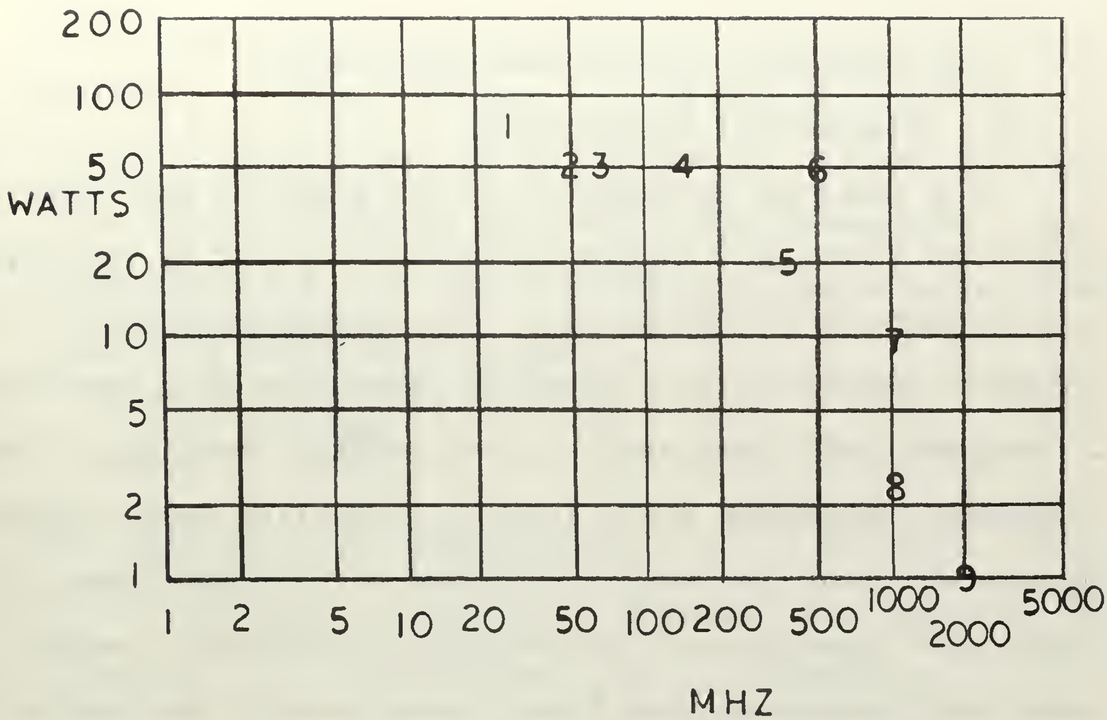

Figure 1. High-frequency, high-power transistor
Class C power output versus frequency.

Table 1. High-frequency, high-power transistors.

Number (above)	Transistor	Manufacturer	Watts @ (class C)	Mhz
1	TA7036	RCA	75	30
2	2N3950	Motorola	50	50
3	2N4130	ITT	50	70
4	2N5214	ITT	50	150
5	TA7036	RCA	20	400
6	2N5178	TRW	50	500
7	S1050	Electronic Comp.	10	1000
8	2N4012	Motorola	2.5	1002
9	TA7003	RCA	1	2000

II. HIGH-FREQUENCY, HIGH-POWER TRANSISTORS

1. Theoretical Limitations.

If a limit is to exist as to the power or gain available from a device, it must ultimately be tied to the physical properties of the device. For transistors, E. O. Johnson[1] related certain physical properties to a theoretical maximum power obtainable, given certain conditions. Considerable idealizing and a lengthy derivation were required by Johnson, but the results obtained offer significant insight into transistors, particularly significant insight into high-frequency, high-power transistors. The derivation is summarized in the following.[2]

For an idealized transistor, we may define a charge-carrier transit-time cutoff frequency,

$$f_T = \frac{1}{2\pi T} ,$$

where T is the average time for a charge moving at an average velocity, the drift velocity, v, to traverse the emitter-collector distance, L.

$$T = \frac{L}{v} .$$

For a given L, T is a minimum for a maximum v.

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1. E. O. Johnson; "Physical Limitations on Frequency and Power Parameters of Transistors", <u>RCA Review</u>, June 1965.

2. Others, notably <u>Electronics</u>, 19 February 1968, have also summarized Johnson's work.

The drift velocity, v, is directly proportional to the emitter-collector voltage, V. V has a maximum, V_{max}, which is the voltage that produces the dielectric breakdown field, $\mathcal{E}$, across the base of the transistor. The magnitude of $\mathcal{E}$ of course depends upon the material from which the transistor is made.

$$\mathcal{E} = \frac{V_{max}}{L}$$

For silicon, $\mathcal{E}$ is approximately 2×10^5 volts/cm. The drift velocity, v, exhibits a maximum, the saturated drift velocity, v_s. For silicon v_s is approximately 6×10^6 cm/sec.[3] For this maximum velocity we then have a minimum transit-time, T_{min}, and a corresponding "maximum" cutoff frequency, f_{Tmax}.

$$T_{min} = \frac{1}{v_s}$$

$$f_{Tmax} = \frac{v_s}{2\pi L}$$

since $\quad V_{max} = \mathcal{E} L,$

we can write, $\quad V_{max}\, f_{T\,max} = \dfrac{\mathcal{E} v_s}{2\pi}, \qquad (1)$

which for silicon is, $\quad V_{max}\, f_{Tmax} \doteq 2 \times 10^{11}$ volts/sec.

Again idealizing, if Q is the total fixed charge in the emitter-collector space, and T is the time for a fixed

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3 v_s is reached at approximately 1/10th of the breakdown field, though the fact is not of significance here.

charge carrier to traverse the emitter-collector distance, L, then the load current can be defined as,

$$I = \frac{Q}{T},$$

and, $$I_{max} = \frac{Q}{T_{min}},\text{[4]}$$

where maximum and minimum conditions apply for V_{max}. Let $Q = C_o V$, where C_o is the "usually quoted value of the collector-base capacitance".

Then, $$\frac{I_{max}}{C_o} = \frac{V}{T_{min}} .$$

Using equation (1) we can write,

$$\frac{I_{max}}{C_o} = \frac{\ell v_s}{2\pi\ f_{Tmax} T_{min}}$$

and since

$$2\pi\ f_{Tmax} T_{min} = 1$$

$$I_{max} = \ell v_s C_o \qquad (2)$$

Combining (1) and (2),

$$V_{max}\ I_{max}\ f_T = (\ell v_s)^2\ \frac{C_o}{2\pi} .\text{[5]} \qquad (3)$$

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4 This idealized condition assumes that all of the base volume is passing charge carriers at the saturated velocity, without base widening. Nothing has been said or implied about achieving this condition, but will be later.

5 f_{Tmax} has been dropped in favor of f_T which corresponds with general use in the literature as the cutoff frequency, and is less redundant.

Departing from Johnson's path, if we let

$$V_{max}\, I_{max} = P_{max} ,$$

$$C_o = \frac{\epsilon A}{L} ,$$

and (again)

$$f_T = \frac{1}{2\pi T} = \frac{v_s}{2\pi L} .$$

Equation (3) becomes,

$$P_{max} = \mathcal{E}^2 v_s \epsilon A, \tag{4}$$

where A is the "effective" base area. By effective, is meant the base area which carries a saturation flow of charge carriers. For silicon, equation (4) becomes,

$$P_{max} \doteq 2.4 \times 10^5 \text{ A watts}, \tag{5}$$

where A is in cm^2 and silicon's relative dielectric constant is approximately 12.[6]

Johnson rearranged equation (3) by letting

$$V_{max}\, I_{max} = P_{max}$$

and,

$$X = \frac{1}{2\pi f_T C_o}$$

to obtain,

$$(P_{max} X)^{\frac{1}{2}} f_T = \frac{\mathcal{E} v_s}{2\pi} \tag{6}$$

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6 W. R. Runyan; Silicon Semiconductor Technology, Texas Instruments Electronics Series, McGraw-Hill,1965,p185.

For silicon, equation (6) becomes,

$$(P_{max}X)^{\frac{1}{2}} f_T \doteq 2 \times 10^{11} \quad \text{volts/sec.}$$

By letting $(P_{max}X)$ be alternately V_{max}^2 and $I_{max}^2 X^2$, we can write,

$$V_{max} \doteq \frac{2 \times 10^{11}}{f_T} \quad \text{volts,} \tag{7}$$

and,

$$I_{max} \doteq \frac{2 \times 10^{11}}{f_T X} \quad \text{amps.} \tag{8}$$

As one should suspect, actual transistor performance is considerably less than the above maxima.

Equation (5) completely ignores what Johnson was interested in demonstrating, but does explicitly demonstrate the area-versus-power concept. Equations (5), (7) and (8) show that for increased power capabilities at a given frequency, a transistor must have increased current-handling capabilities, which requires its effective base area to be larger. For an increase in current capability, the impedance level must be lowered.

High-frequency, high-power transistors are exclusively of silicon construction. The other commonly used semi-conductor material, germanium, is excluded almost entirely from high-frequency, high-power applications because of its lower $\mathcal{E}v_s$ product ($\frac{1}{2}$ that of silicon), and because intrinsic conduction temperature for germanium is lower than that of silicon; hence a larger heat sink would be required. See part 4, this section.

Equations (4) and (6) suggest a search for materials

with a higher $\mathcal{E}v_s$ product be conducted, as a possibility for future higher-frequency, higher-power transistors. (see part 4 this section.)

2. Physical Design.

Considering a transistor as a silicon wafer, with a planar emitter and collector having been diffused in from opposite sides of the wafer, and maximum base thickness being determined by frequency considerations (as discussed previously), the seemingly obvious method to increase the power of the transistor is to make the wafer, including the emitter and collector, larger in the dimension parallel to the junctions.

Thus it would seem that the size of the crystal wafer it is possible to grow would be the limiting factor in the power obtainable from a transistor. N. H. Fletcher[7] demonstrated the invalidity of this approach after showing that a thin base is desirable for high base-transit efficiency. He demonstrated that the portions of the base removed from the base contact (conductor metal) pass little current. The "cross-base" current reduces the electrical potential of those remote portions of the base. Thus in a planar transistor, only the portion of the base near the emitter edge is effective in passing current, and the larger the base current, the more pronounced is this effect. Fletcher proposed (in 1955) a very long thin emitter for a high-frequency, high-power transistor. An effective emitter for

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7 N. H. Fletcher; "Some Aspects of the Design of Power Transistors", Proceedings of the IRE, Vol 43.2, May 1955, pp 551-559.

a high-frequency, high-power transistor should have a large amount of "edge" for a given area, termed universally as "edge-area ratio" or "periphery-area ratio", or more recently just as a high "periphery".

The achievement of a high periphery lead from a large circular emitter design to long thin emitter designs, to comb and multiple emitter designs. The one problem that multiple emitters (called overlay transistors) and comb structured emitters (called interdigitated transistors) solved was to put a large area of emitter in close electrical contact with an equipotential base. That is, a large amount of base-emitter junction is electrically close to the base lead.

Construction of both overlay and interdigitated transistors involves the use of precisely constructed, precisely located photo-masks, as only with thin emitter structures in close proximity to the base lead(s) will a large percentage of the base have anywhere close to the number of injected charges that it can transport. (Recall that the theoretical power discussion supposed a saturated flow of charges in the base.) Overlay and interdigitated transistors could be thought of as large base area transistors that use the base effectively.

The Overlay Transistor.[8]

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8 D. R. Carley, P. L. McGeough, and J. F. O'Brien; "The Overlay Transistor, Part I: New Geometry Boosts Power", Electronics, 23 August 1965, pp 71-77.

The name overlay refers to the fact that the emitter metal overlays the base. The basic steps involved in building an overlay transistor are (npn assumed);

a. A masking operation defines the base region on a n-doped silicon substrate. A light p-doping is evaporated onto and baked (diffused) into the surface. The surface is oxidized during this operation.

b. After deoxidization of the surface by a second mask, p+ regions (grids) are evaporated onto and baked into the chip, causing another oxidization of the surface. These are the highly conductive regions that will carry the cross-base current with little cross-base voltage drop.

c. The emitter mask is then positioned so that the chip is deoxidized in the center of each p+ base grid. Into these deoxidized areas in the center of the base p+ grids, the emitter areas are created. One hundred or more emitter areas are not uncommon. The emitters are created by evaporating sufficient n material (sufficient to convert the overall doping level back to donor or n) onto the exposed areas and diffusing it into the chip, with subsequent oxidization of the surface.

d. All emitters must now be connected to a common metal connection. Removal of the oxide above the emitters to allow connection in common, without contact with the base region, is accomplished by a precision located mask. Then an overlay of aluminum is evaporated onto the chip, connecting all emitters in common. The aluminum overlays

approximately one half of the p+ base grid region. At the same time an electrically separate deposit of aluminum connects all of the exposed p+ base regions in common.

Steps a. through d. above correspond to figures 2a through 2d following, for a hypothetical overlay transistor. Figure 3 shows a perspective view of the same hypthetical transistor.

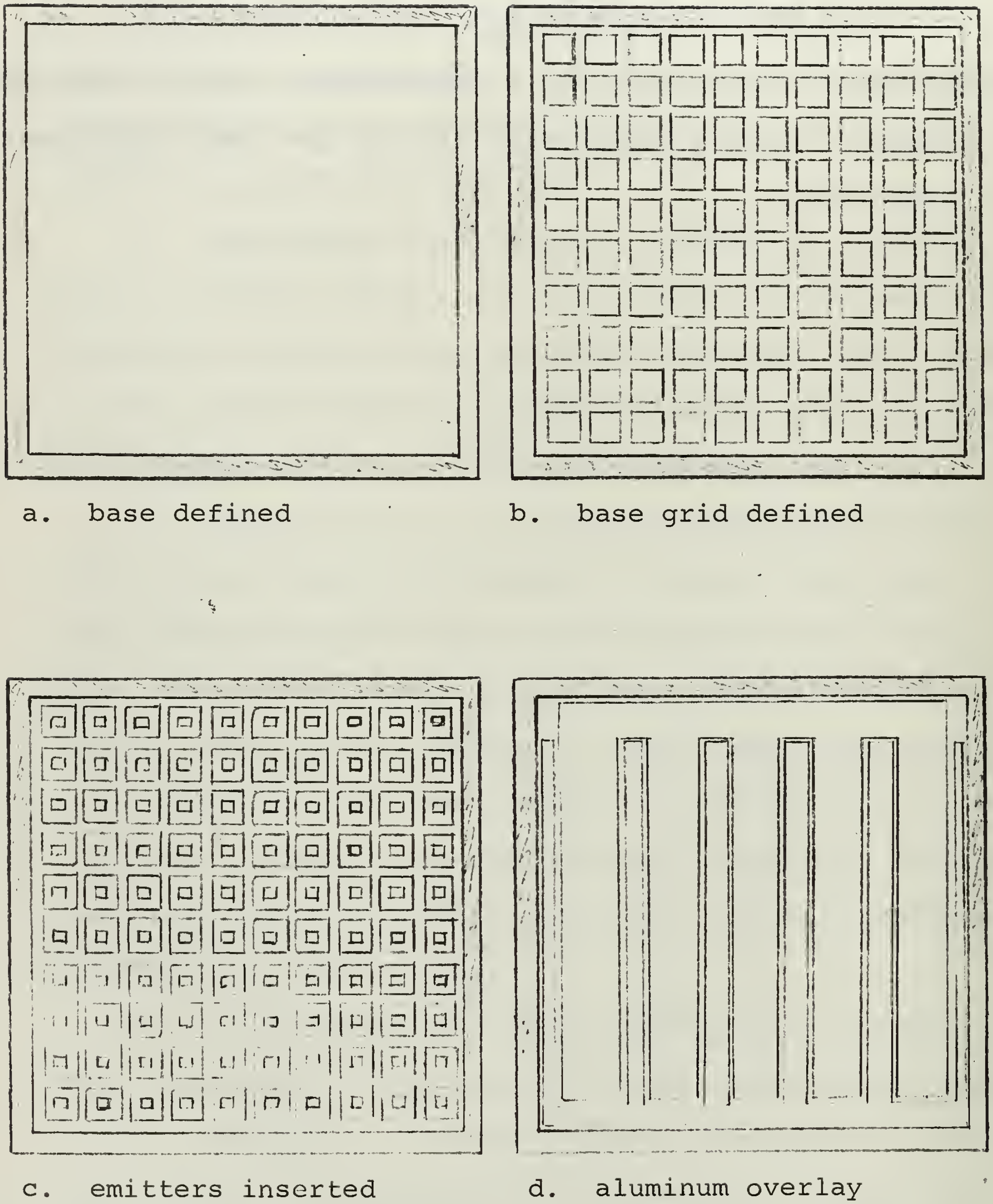

Figure 2. A hypothetical overlay transistor in various stages of completion.

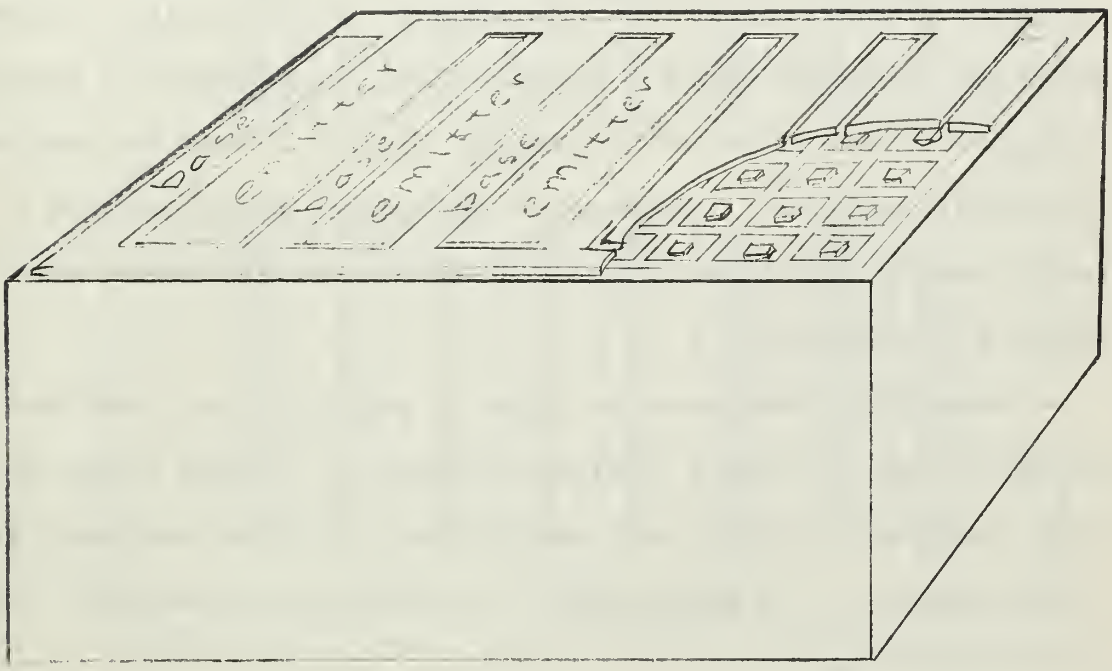

Figure 3. Perspective view of hypothetical overlay transistor.

The Interdigitated Transistor.[9, 10]

The interdigitated transistor is different from the overlay transistor only in the physical layout of its emitters and p+ base region. In the interdigitated transistor, the emitters and p+ base regions are alternate long thin bars, set in the base with the long thin bars connected at opposite ends. Both the p+ base and the emitter regions resemble combs with teeth (or fingers) interlocking, hence the name interdigitated. Figure 4 following shows a hypothetical interdigitated transistor (without base and emitter metallization). Comparable photomask sharpness and postitioning skills are required with interdigitated and overlay transistors.

Generally reference to type of construction need not be made from a circuit design standpoint, though often design engineers will do so, particularly if the engineer is in the employ of a manufacturer that manufactures only one of the two types.

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9 New Components; "Power Transistors: The Jump to 50 Watts", <u>Electronics</u>, Vol 40, No. 17, 21, August 1967, pp 150-152.

10 C. Fromberg; "The Capabilities of UHF Power Transistors", Ministry of Aviation, Farnborough Hants, (U. S. Government Nr. AD 644 435), June 1966, p3.

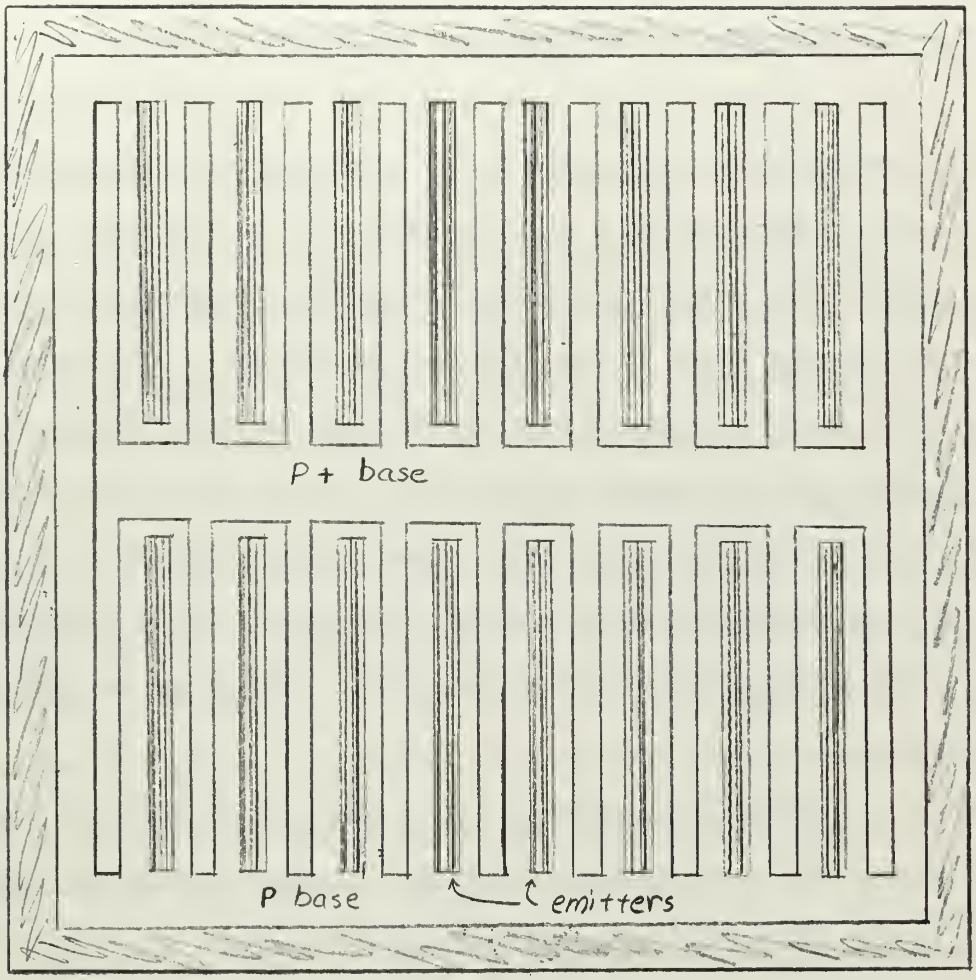

Figure 4. A hypothetical interdigitated transistor, npn.

3. Some Notes on Second Breakdown.[11]

Most articles on the subject of second breakdown[12] meticulously state that second breakdown is not just a high-frequency, high-power transistor phenomenon, but can occur to any transistor and that the phenomenon of second breakdown is not completely understood. Nevertheless, second breakdown did not receive any widespread attention until the advent of high-frequency, high-power transistors, and since then it has caused problems. An example is a 1963 Radio Corporation of America attempt to build a 100-watt, 25 to 50-Mhz transistor, ending in "catastrophic failure phenomenon known as 'second breakdown'......". Nearly all reports of more recent date concerning high-frequency, high-power transistors pay their respects to second breakdown.

Second breakdown is a condition where the output impedance of a transistor changes almost instantaneously from a large value to a small limiting value. It may be distinguished from normal transistor operation by the fact that

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11 H. A. Schafft and J. C. French; "Studies of Second breakdown in Junction Devices",National Bureau of Standards, Dec. 1965. This is the final report on a series of studies of second breakdown, which not only provides much useful information, but also lists some 43 other second breakdown references.

12 The term second breakdown is somewhat unfortunate, as the phenomenon is not necessarily related to a first breakdown. It is more aptly described as a second kind of breakdown, that is not related to a voltage above the breakdown voltage of the device.

once it occurs, the base no longer controls normal collector characteristics.[13] Second breakdown can occur when the device is otherwise within its operating limits and when it is either forward or reverse biased (conducting or cutoff).

Second breakdown can best be described briefly as localized thermal runaway.[14] Some measure of protection against second breakdown is afforded by placing emitter resistance within the transistor, either at each emitter site for overlay transistors,[15] or in the connected end of each finger of the emitter comb of interdigitated transistors. An analogy to this would be to consider each emitter spot (overlay type) or each finger (interdigitated type) as a separate transistor, all of them to be connected in a parallel common-emitter circuit. Emitter resistors are used to evenly distribute the current and prevent individual overloading with subsequent thermal runaway.

To insure complete protection against the possibility of second breakdown, a circuit designer must conduct a series of tests on the particular transistors to be used,

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13 P. Schiff; "Second Breakdown in Transistors", Radio Corp. of America Application Note SMA-30, 1964, p.l.

14 C. Turner; "Carl Turner of RCA Speaks Out on Second Breakdown", <u>EEE</u>, Vol.15,Nr.7, July 1967, p.4.

15 R. Rosenzweig and Z. F. Chang; "Development of a 50-Watt, 80-MC Linear Amplifier Transistor", Radio Corp. of America, March 1966, pp8-12.

under the conditions of use, since manufacturer's specification sheets are necessarily (cost wise) incomplete with regard to all conditions which might cause second breakdown. References 12, 13 and 14 (see previous pages) offer some guidance on testing and circuitry. Some general guidelines are:

a. Under pulsed conditions, transistors are less susceptible to second breakdown for short pulses.

b. To avoid second breakdown, the circuit designer should select the lowest-frequency, lowest-voltage transistor possible.

c. Any coils (or RFCs in the collector circuit should have as little inductance as possible.

4. Advancing State-of-the-Art.

Use of internal emitter-resistor stabilization and internal grounding of the emitter lead to the case have both helped bring the state of the art to its present level.[16] The internal emitter resistors, sometimes referred to as ballasting resistors,[17] required on the larger of the multiple-emitter devices, serve to prevent local thermal runaway (second breakdown) or to increase the size of the safe operating area. The safe operating area is that area on the collector current versus collector voltage plane where the transistor is relatively safe from second breakdown. Internal grounding of the emitter lead to the case reduces the emitter lead inductance, which is vital for common-emitter circuits.

At frequencies above a few gigahertz, materials other than silicon (materials with a higher $\mathcal{E}v_s$ product) have significant advantages, sufficient to warrent their use even at the expense of developing the technology required. GaAs and other materials from the third and fifth valence columns of the periodic chart of the elements are being actively investigated.[18] Even amorphous materials, such as

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16 J. G. Tatum; "VHF/UHF Power Transistor Amplifier Design", International Telephone and Telegraph Corp. Application Note, 1967, p1.

17 Radio Corp. of America; "Transistor, VHF, Silicon, Power, Linear, 30Mhz, 100Watts PEP", December 1966, p.19.

18 M. B. Leeds and others; "A Chapter Ends--- But the Story Continues". Electronics, Vol. 41, Nr.4, 19 Feb.1968,p130.

chalcogenide glass are being considered.[19]

Of a more evolutionary nature, reference 20 describes a mesh emitter transistor (MET). The visible difference between MET and the overlay transistor is that in the MET the emitter is a mesh or grid enclosing small base n+ regions that are connected in parallel by metallization. Proponents of METs claim that for a given state of the mask-making art, MET's have a higher periphery than either overlay or interdigitated transistors. The present state of the art is that overlay transistors provide a saturated drift-velocity flow of charge carriers over approximately 70 percent of the base area, so the improvement expected from a MET will not be greater than a factor of 100/70.[21] Remaining to be worked out in METs are adequate emitter resistors.

Class-B or class-AB biasing is particularly difficult to maintain under varying signal and temperature conditions. The reasons for class-B or class-AB biasing will be discussed in detail in section III. Bypassed emitter resistors which would nearly prevent bias point shifting for small-power transistors represent a significant loss of power and

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19 Science and the Citizen; "The Glass Switch", <u>Scientific American</u>, Vol.218,Nr.2,February 1968, p52.

20 M. Fukuta and others, "Mesh Emitter Transistor", Letters Section, <u>Proceedings of the IEEE</u>, April 1968,pp742-3.

21 E. O. Johnson; "Physical Limitations on Frequency and Power Parameters on Transistors", <u>RCA Review</u>, June 1965, p165.

efficiency for high-power transistors. The bypassed emitter resistor and the bias-voltage source can be thought of as a constant current biasing scheme. Biasing with a constant current source for a high-power transistor would solve the bias-point change-with-temperature problem, but is unsatisfactory due to the rectifying action of the base-emitter junction, which for different drive levels would allow the bias point to shift. (Nor does constant-current biasing, even with the aid of external emitter resistance materially aid in keeping the current distribution uniform within the large area device; this function is accomplished only by internal emitter resistance and uniform geometry.) A constant-voltage bias would provide the correct bias for all drive levels, but would allow bias-point drift with temperature change. (Note that temperature change in the transistor can be due solely to differing amounts of power dissipated in the transistor, rather than a change in environmental condition.) A good compromise to the above dilemma is a biasing amplifier capable of supplying a large current at a voltage that is referenced (and equal) to a slightly forward biased pn junction voltage at the same temperature as the transistor. The temperature of the pn junction needs to be the same temperature as the transistor, as a pn junction's voltage decreases at a rate of approxi-

mately 2 millivolts per celsius.[22] Radio Corporation of America's device TA2758 aids in accomplishing the above by poviding a diode and a high-frequency, high-power transistor on the same chip. Not only is the thermal time lag minimized, but doping levels are identical. An external unity-voltage, current amplifier provides the current gain needed to provide a constant class-AB bias with temperature stability.[23]

Providing a low thermal gradient heat path from the active chip to a suitable heat sink has always been of critical design concern in high-frequency, high-power transistor fabrication. As the power levels get higher and higher, it becomes more critical, and may well dictate chip size and placement. In at least one case it already has.[24]

To digress briefly, the basic reason that the temperature must be kept within a certain bound is that a transistor's operation is dependent upon non-intrinsic conduction. The onset of intrinsic conduction, occurring with high temperature, swamps the transistor action.[25] Thus a

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22 The new correct (but perhaps unfamiliar) notation for temperature eliminates the use of degree (^{o}) and of course centegrade has been replaced with celsius.

23 Radio Corporation of America; RCA Silicon Power Circuits Manual, March 1967, pp300-303.

24 "A 50-Watt 500Mhz Transistor is Announced", Electronic Capabilities, Winter 1967/68.

25 L. V. Azaroff and J. J. Brophy; Electronic Processes in Materials, McGraw-Hill, 1962, pp12-29.

transistor's temperature range of usefulness is below the temperature of intrinsic conduction (approximately 200C for silicon and 85C for germanium). Thermal instability manifests itself in increasing leakage current (collector to base), increasing forward conduction current (emitter to base), increasing forward conduction current (emitter to base, approximately 8 percent per celsius), and increasing non-uniformity of current distribution in large area devices, with increasing temperature.[26] A second-order effect is a decrease in f_T at higher temperatures.[27]

A commonly defined constant for a high-frequency, high-power transistor is a thermal resistance, defined as the temperature difference between the junction temperature and the transistor's case, per watt of power being dissipated. However, the thermal resistance is not a constant. The measured value of thermal resistance becomes larger for higher current levels due to higher current concentations in certain spots within the transistor. Note that when a small area of a transistor tends toward thermal runaway, its conductance increases and more current is funneled into that area.[28] See part 3 in this section.

The most common thermal resistance among the present

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26 J. G. Tatum; VHF/UHF Power Transistor Amplifier Design", ITT Application Note, 1967, pp12-13.

27 L. P. Hunter; Handbook of Semiconductor Electronics, 2nd Edition, McGraw-Hill, 1962, pp12-29.

28 J. G. Tatum; "Microwave Transistor", ITT Application Note, 1967, pp18-19.

high-frequency, high-power transistors is about 2.5C/watt. This value of thermal resistance itself limits the power dissipation of a high-frequency, high-power transistor with an infinite heat sink to about 70 watts, assuming maximum junction temperature of 200C and a heat sink temperature of 25C. Assuming an efficiency of 60 percent for class AB operation, the high-frequency, high-power transistor cannot deliver more than 100 watts peak. Some reduction in the 2.5C/watt figure will result from a larger active area, but not nearly the magnitude needed for larger power devices.[29]

The heat sinks needed for today's high-frequency, high-power transistors, due to the large thermal resistance, make a transistor power amplifier about equal in weight to a vacuum-tube power amplifier when the weight of the tube's filament transformer is included, and probably heavier than battery-powered sets where filament transformers are not needed.[30] For device safety purposes, particualrly where a distinct possibility of a sudden complete mismatch of output circuit exists, as is the case in military field radios, the device should be capable of dissipating all the power delivered to it, or approximately twice its output rating.[31]

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29 RCA; "Transistor, VHF, Silicon, Power, Linear, 30Mhz, 100-Watts PEP", December 1966, p14.

30 RCA; "Solid State, Broadband, 100-Watts, 25 to 50Mhz Transmitter", August 1963, p.16.

31 G. Ewing; Lecture on Solid State Circuit Design, EE3263, Naval Postgraduate School, Term 4, 1967-68.

A lighter, more effective heat sink is needed. With a heat pipe, either as an integral part of a high-frequency, high-power transistor or, less effectively, attached as an external heat sink, a net reduction in the weight of and an improvement in the performance of a transistorized power amplifier may be possible. A heat pipe is essentially a closed, evacuated chamber whose inside walls are lined with a capillary structure, or wick, that is saturated with a volatile fluid.[32] By a combination of vapor heat transfer and capillary action the heat pipe is capable of transporting large quantities of heat energy along its length with a very small thermal gradient, thus creating a structure that has a thermal conductivity orders of magnitude greater than any known material.[33 & 34] Heat pipes contain no moving parts, other than the liquid/gas, and have a lengthy mean time before failure. Heat pipes may be constructed for operation at nearly any temperature range and in some cases may be constructed entirely from electrically insulating materials.[35]

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32 G. Y. Eastman; "The Heat Pipe", Scientific American, Volume 218, Nr.5, May 1968, pp 38-46.

33 G. M. Grover, T. P. Cotter and G. F. Erickson; "Structures for Very High Thermal Conductance", Journal of Applied Physics, Volume 35,Nr.6, June 1964.pp1990-1991.

34 K. T. Feldman, Jr. and G. H. Whiting; "The Heat Pipe", Mechanical Engineering, Volume 89, Nr.2, February 1967, pp30-33.

35 G. Y. Eastman; "The Heat Pipe", Scientific American, Volume 218, Nr. 5, May 1968, p46.

Normal overlay and interdigitated transistors are "built" onto one side of a thin chip of silicon (planar construction). On that one side electrical connections are made to the emitter and base. The heat energy which must be dissipated from the transistor chip is created at the collector-base junction, very near that one side. The main path for the dissipation of the heat energy is across the chip and into the metal case, often the path is also through an electrically isolating layer of beryllium oxide. A little heat energy is dissipated from the near side, through the leads. Reference 36 presents a Radio Corporation of America investigation of a device that dissipated the heat energy in both directions, with a subsequent halving of the otherwise best obtainable thermal resistance. The technique was essentially to construct the emitters (a multi-emitter device was constructed, similar to an overlay transistor) and base grid on one silicon chip, complete with emitter resistors. The second chip contained the collector and a thin base. After the individual construction and surface polishing, the two chips were fused together in a manner that allowed the emitter-base junction to drift slightly away from the physical chip junction. Electrical connections were made at the edges and good thermal contacts were

36 H. W. Becke and D. Stolnitz; "Investigation of Laminated Structures for Radio Frequency Applications", Radio Corp. of America, September 1967.

made on both sides. The results were very encouraging, the thermal resistance being halved. However more developmental work is needed due to several problems, which are discussed in the report.

5. Characterization for Power Amplifier Purposes.

Characterization of a transistor as a grouping of circuit elements is of little interest if the resulting characterization, no matter how accurate, is of no use for the purpose of circuit analysis and design. Practically this places a limit on the complexity of the characterization. Attempts to progress a step at a time from a three-dimensional physical device (such as figure 3) to a two-dimensional schematic become very complex. Hybrid-pi and T equivalent circuits are the result of compromising accurate characterization and simplicity. The characterizations such as figures 5 and 6 represent rather well the small-signal behavior of a common-emitter configured transistor

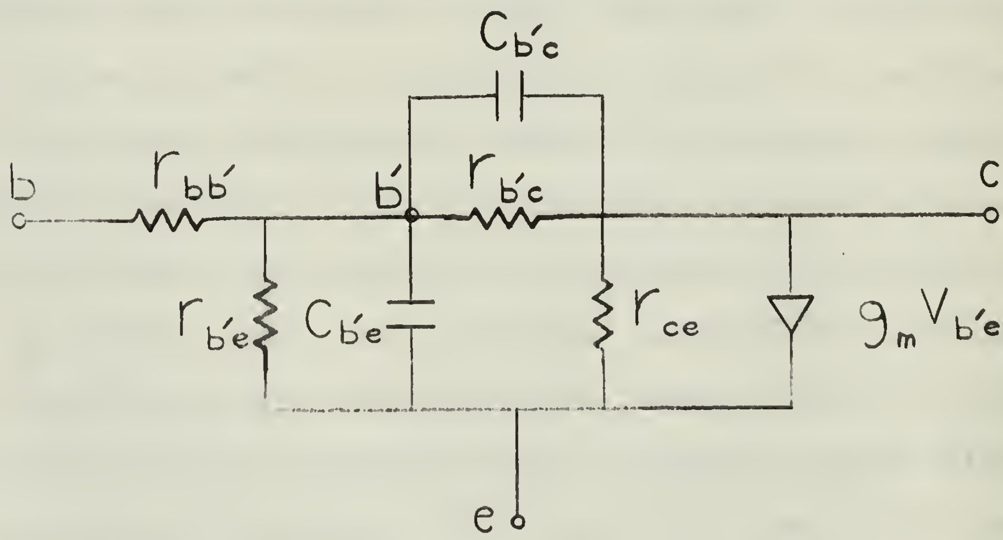

Figure 5. Hybrid-pi characterization of a common-emitter high-frequency, high-power transistor.

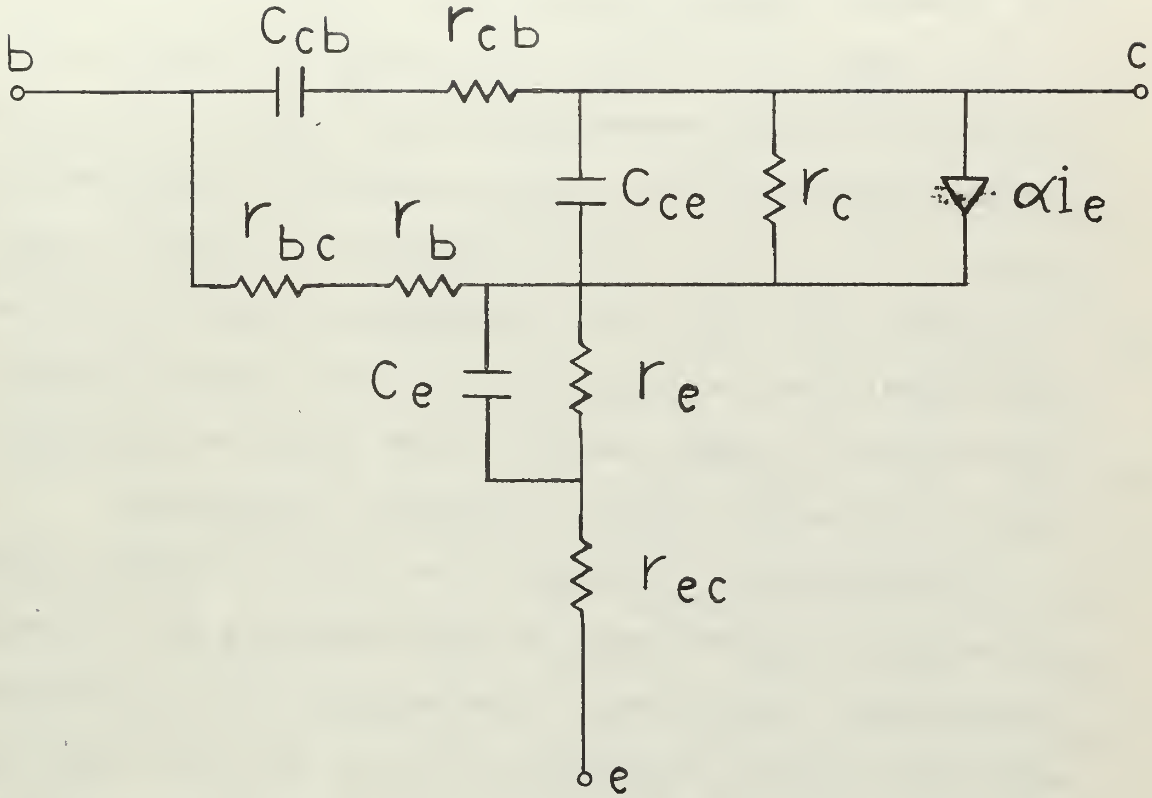

Figure 6. T equivalent circuit characterization of a common-emitter high-frequency, high-power transistor.

over its useful frequency range.[37, 38 & 39] The complexity of measurement techniques required to determine values has, however, reduced the popularity with circuit designers.

The need for accurate (and perhaps more importantly, the need for easily measurable) quantities has led to the common acceptance of the various parameter equivalent circuits. A, b, g, h, y and z parameters are found in the literature. The six sets of parameters represent (if nothing else) the six different ways that input and output currents and voltages can be related using two equations with two dependent and two independent parameters.

H-parameters (for hybrid) are widely accepted, because of the ease of measurement of the parameters and a certain "naturalness" about them. Y-parameters have an important advantage at high frequencies, in that they are short-circuit parameters, and can be more accurately measured than parameters which require an open circuit. At high frequencies valid open-circuit measurements are difficult to achieve, due to the effects of stray capacitance on measured parameter values.[40] At high frequencies the parameters

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37 R. Milton; "Design of Large-Signal VHF Transistor Power Amplifiers", Radio Corporation of America Application Note SMA-36, July 1964, pp.1-2.

38 L. P. Hunter; Handbook of Semiconductor Electronics, McGraw-Hill, 1962, Chapter 11, page 16.

39 J. G. Tatum; "VHF/UHF Power Transistor Amplifier Design" ITT Application Note, 1967, p5.

40 F. C. Fitchen; Transistor Circuit Analysis and Design, D. Van Nostrand, 1960, pp 91-95.

generally are not constant with frequency and hence cannot be simply stated but must be plotted versus frequency. Also at any given frequency the parameters generally are not constant with a change in bias point.

As will be discussed in section III, part 1, it is usually necessary that the final stages of a power amplifier operate in or near class B operation, for linear power amplification. To achieve a high output power it is also necessary for the transistors to handle large signals. The transistor will be in cutoff, active and (at least near) saturation regions during each large-signal cycle. This is the worst possible mode of operation from the standpoint of attempting to characterize the transistor by an equivalent circuit composed of linear elements. Small-signal parameters are also inaccurate to characterize large-signal operations, though for some particular case this inaccurate characterization can be modified to be more accurate.

At least two transistor manufacturers (Motorola, Inc. and International Telephone and Telegraph Corp.) presently provide large-signal input and output impedance data to aid the circuit designer.[41 & 42] The large-signal impedance data provided by the manufacturers is for transistors oper-

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41 R. C. Hejhall; "Systemizing RF Power Amplifier Design", Motorola Application Note 282, Motorola Semiconductor Products, Inc. 1967.

42 J. C. Tatum; "VHF/UHF Power Transistor Amplifier Design", ITT Application Note, 1967. p32.

ating in the common-emitter configuration, biased class C. The data is of very little or no use for other configurations. The data seems to be nearly exact for class B use, and at least it is better than that obtained from small-signal measurements and modified by rules of thumb.

The equivalent circuit suggested from the data supplied by Motorola is shown as figure 7, a below, where the power generator is of an unknown character and is included solely as a reminder of where the output power is derived.

The most useful equivalent circuit for a class-B, high-frequency, high-power transistor is with input series elements and parallel output elements, as suggested by ITT and shown as figure 7, b below, where again the power generator is of unknown character and is included for completeness. Methods of obtaining impedance data shall be discussed in section III, parts 3 and 4.

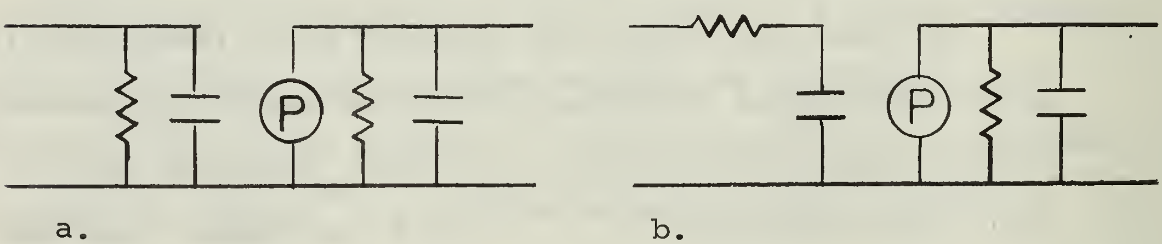

Figure 7. Best characterizations of high-frequency high-power transistors for class-B power amplifier.

III. LINEAR POWER AMPLIFIER DESIGN

1. General Aspects.

A high-frequency power amplifier is considered to be linear if it is capable of producing at its output an amplification of signals provided to its input, such that the relative phase and magnitude of the various frequency signals at its input and output are the same and no additional signals are produced by the amplifier. For general communication purposes, the amplifier should be capable of doing this over a range of approximately 20 kilohertz to each side of a center (carrier) frequency.

The tests for linearity and measures of relative linearity are a ratio (in db) test between the power levels of two applied test signals at the input and output of the amplifier, and an intermodulation distortion test, which is a ratio (in db) between the cross harmonics and the two output signals when the two signals are of equal power at the input.

The ideal linear amplifier is of course class A[1], but the low efficiency of class A amplifiers precludes them from consideration as power amplifiers in many power amplifier situations. Discussion of class A amplifiers and class A large-signal amplifiers is not irrelevant, however,

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1 A class A amplifier is biased so that collector current is flowing constantly and an input signal alters the amount of collector current in a linear fashion. Maximum theoretical efficiency is 50 percent, with practical efficiencies of about 25 percent.

due to their usual employment as a driver for the final stages.

The highest efficiency power amplifier is a class C amplifier.[2] Class C amplifiers contain much distortion in the form of harmonic components. In situations where at any instant only a single-frequency signal output is desired, a highly selective circuit may be used to drastically reduce the higher frequency components of the output signal. A class C amplifier is used for highest possible efficiency. A class C biased transistor is not a linear amplifier, and cannot be used in a linear power amplifier as the discussion below will demonstrate.

For a single input frequency, a class C biased transistor has at its output the original input frequency plus (in decreasing magnitudes) all harmonics. A selective output circuit essentially eliminates the harmonics in this case. With two input frequencies, of nearly equal frequencies, again all harmonics of both frequencies will be present at the transistors output, but unfortunately so will all "cross" or "modulation" products of the harmonics of the two frequencies. Of primary importance are the cross products of the second and third harmonics, since their difference frequencies will be near the original two fre-

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2 A class C amplifier is biased so that collector current flows only when the input signal voltage is near its peak. Maximum theoretical efficiency is 100 percent, with practical efficiencies as high as 85 percent.

quencies and any subsequent selective circuitry will not eliminate them. Figure 8 following shows sketches of signals in the frequency domain for a single frequency and two frequencies. The terms for the two frequency case can be obtained from a convolution technique,[3] or from listing all harmonics of each frequency and adding all possible sum and difference combinations of the two listings, or from a complicated and uninteresting Fourier series. A linear amplifier must in reality pass not just two but a multitude of different frequency signals without creating any cross products. It is reasonable to assume that an amplifier that is without cross products for two input signals will be without cross products for a multitude of input signals.

Class A operation is unacceptable from an efficiency standpoint; class C operation has been eliminated from a linearity standpoint. Somewhere in between the two is the only other possibility. Class B operation is usually defined as zero bias voltage, or that biasing which allows conduction during 180 degrees or half of each cycle. Class AB biasing is with a slight forward bias voltage, for conduction just slightly more than 180 degrees. The reason for class AB biasing is discussed below.

Transistor linearity is governed predominantly by two

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3 M. Schwartz: Information Transmission, Modulation, and Noise, McGraw-Hill, 1959, pp 64-73.

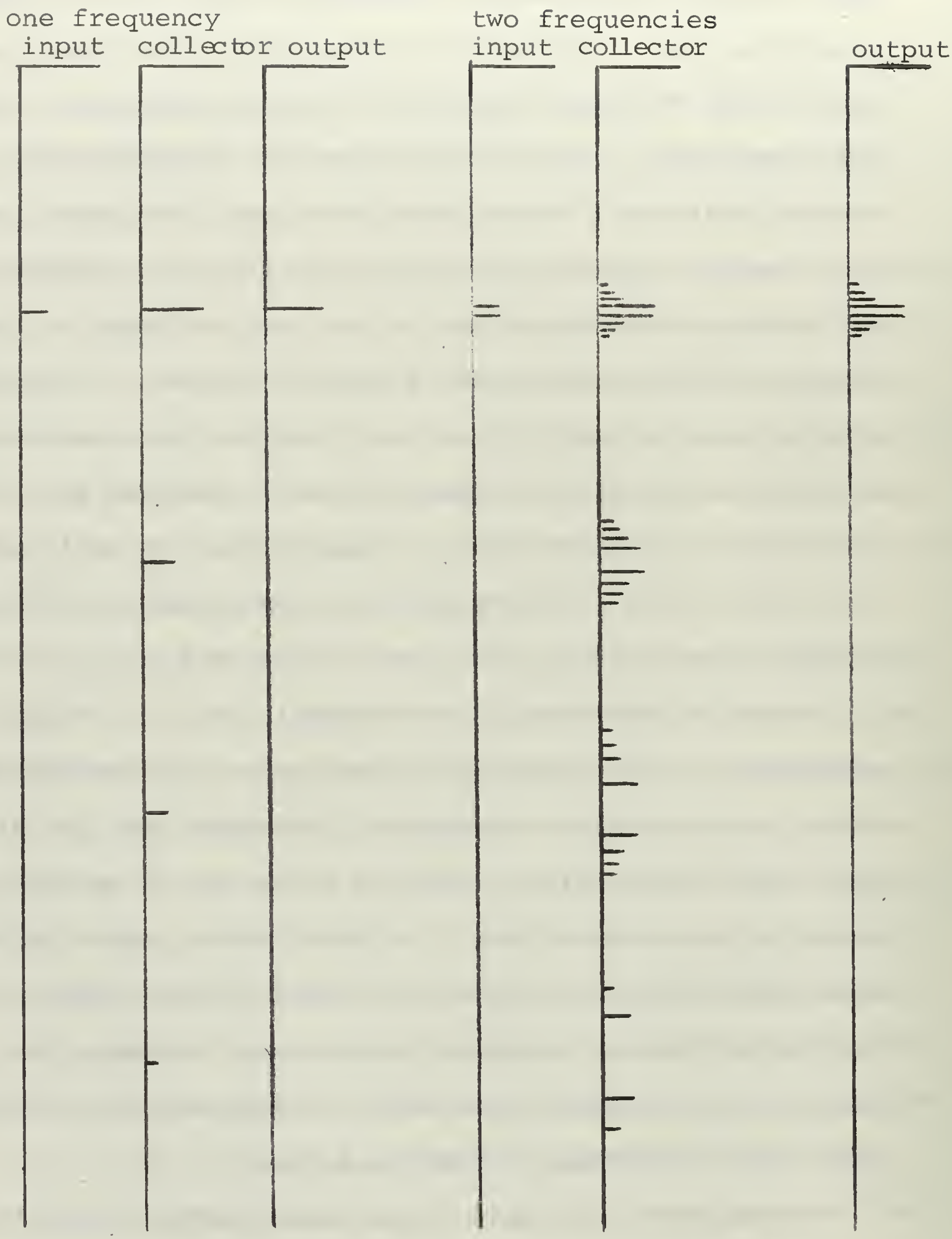

Figure 8. Class C amplifier frequency plots for one and two input frequencies.

factors. The first factor is low-level distortion, caused by the inherent non-linearity of the base-emitter (diode) junction. The second factor is the "peak-current limiting" by the peak current-handling capability of the transistor.[4] Essentially the current gain of the transistor varies near cutoff and saturation, which is to be expected. How near saturation the non-linearity in the transistor's current gain exists is subject to some control, and will be mentioned again later in this section. Large-signal operation at high frequencies with large area (high-frequency, high-power) transistors does not allow an ideal class B (or class AB) operation. The problem is caused partly by large values of collector-to-base capacitance, C_{cb}, and partly by emitter lead inductance and high current concentrations within the transistor.[5] For the above reasons, during the remainder of this thesis, class B operation will be assumed to be a broad enough definition to include class B and class AB biasing.

It remains to be shown that a transistor biased at or near cutoff can be incorporated easily into a linear amplifier. For that purpose the previous class C example (figure 8) and the following examples (figures 9, 10 and 11)

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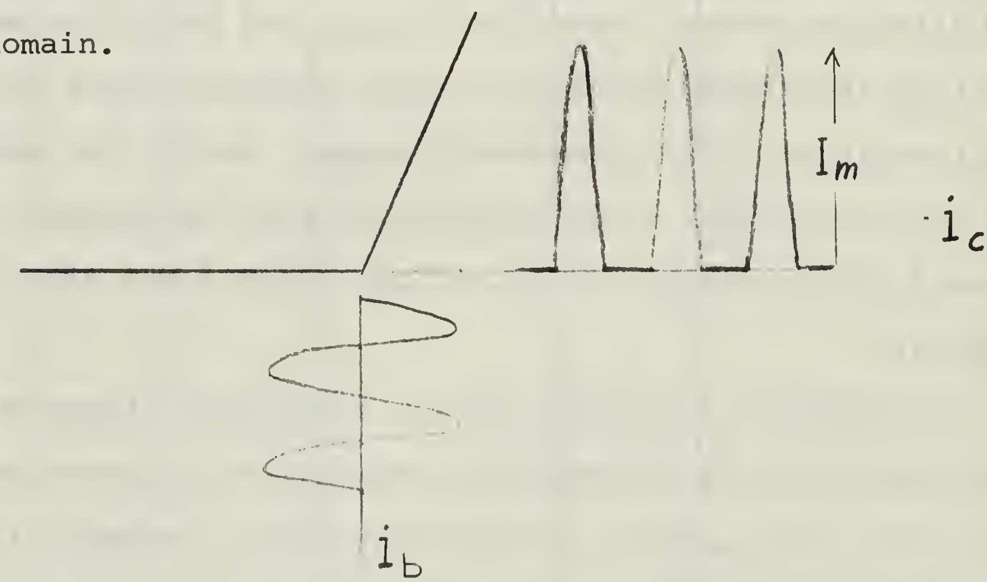
4 Radio Corporation of America; "Transistor, VHF, Silicon, Power, Linear, 30Mhz, 100-Watts PEP", Dec. 1966, p.7.

are given. Notice that in the class C, two-frequency example an odd harmonic is involved in each distortion term. This will be true as long as the bandwidth is much smaller than the operating frequency. Hence a transfer-characteristic curve that created no odd harmonics would be without intermodulation distortion. If a transfer charactersitic were represented by two straight-line segments, and the bias point were at the intersection of the two straight line segments, representing class B biasing, no odd harmonics would be created. Figure 9, below, represents such an idealized situation, along with the Fourier series. Figure 10, shows one-and two-input frequency cases in the frequency domain.



$$i_c(t) = \frac{I_m}{\pi} + \frac{I_m}{2}\text{Sin}\ \omega t - \frac{2I_m}{3\pi}\text{Cos}\ 2\omega t - \frac{2I_m}{15\pi}\text{Cos}\ 4\omega t - \frac{2I_m}{35\pi}\text{Cos}\ 6\omega t \ldots$$

Figure 9. An idealized class-B bias, straight-line idealization.

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5 J. G. Tatum; VHF/UHF Power Transistor Amplifier Design", ITT Application Note, 1967, p.2.

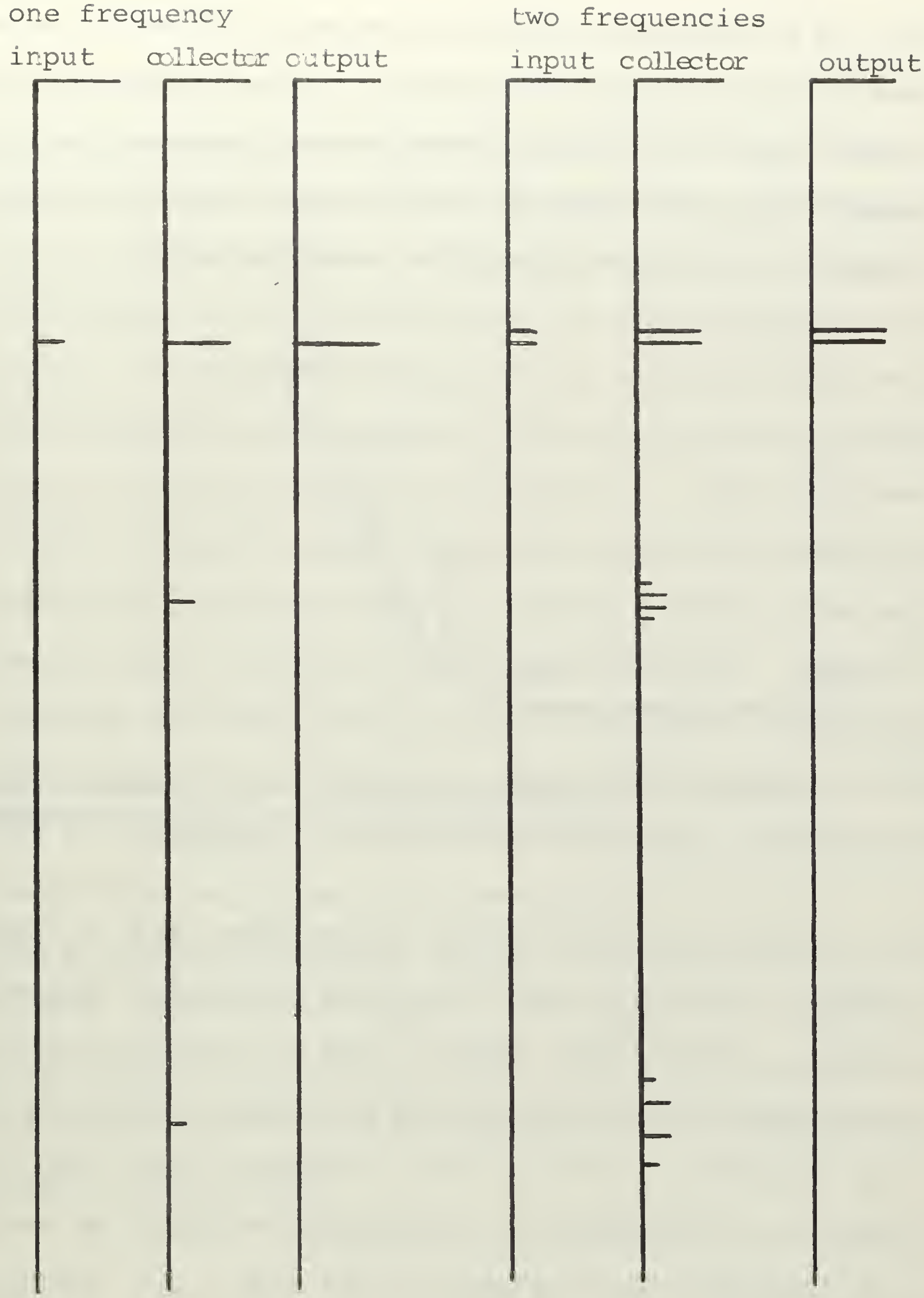

Figure 10. Class B amplifier frequency plots for one and two input frequencies, straight-line idealization.

A better approximation to existing transfer characteristics, though still idealized, is two straight-line segments joined by a second-order curve, connected in such a manner that the slopes are zero at the lower connection and equal at the upper connection, as shown below.

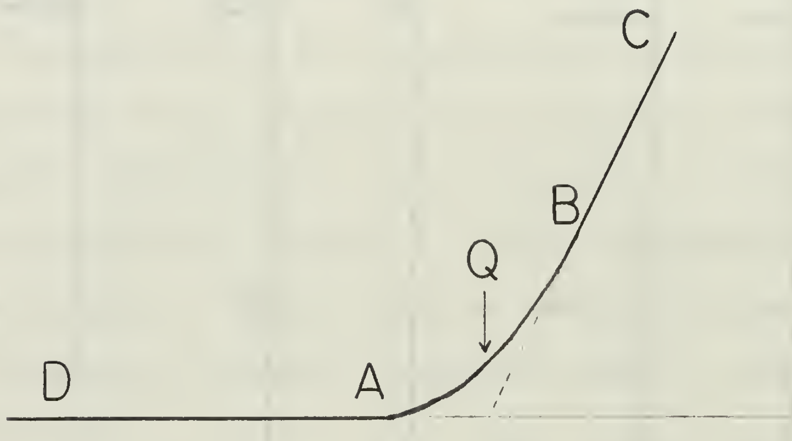

Figure 11. Transfer characteristic, second-order curve connecting straight-line segments.

Point Q is the point of proper bias, and is "projected cutoff". Points A and B are equal horizontal distances from point Q. Small input signals that are located entirely on the second-order curve produce an output containing the input frequency, twice the input frequency and direct current. Hence no intermodulation distortion results from more than one frequency being present at one time. For larger input signals, the analysis becomes slightly more complicated. For a rigorous solution a point-by-point graphical analysis of the time waveforms could be made (Chaffee 11 point analysis, or a modification thereof). Such an example will not

be included here, as a plausibility argument will enable one to visualize the result. Notice that when the input signal becomes larger than the distance QA (and QB) it enters a linear region in both (positive and negative) directions at the same time, with no change of gain (slope), hence there is no change in gain of the fundamental component.[6] No intermodulation distortion results from either the straight-line segments or the second-order curvature as long as the bias point is located at Q. This requires slight forward biasing. Of course on actual transfer characteristic will not display perfect straight-line and second-order curvatures properly positioned. Current crowding at high-current levels is the main cause of deviation from the straight-line segment BC. Current crowding and the subsequent transfer characteristic curvature can be reduced by selecting a low-beta device.[7]
Any additional curvature yields intermodulation distortion. In some cases high-frequency, high-power transistors have been designed specifically for a low beta to improve class-B biased linearity and are specifically recommended for single-sideband use.

A low-beta transistor in a class B or class C common-emitter configuration also has the following advantages over

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6 W. B. Bruene; "Linear Power Amplifier Design", Proceedings of the IRE, December 1956, pp. 1754-1759.

7 J. G. Tatum; "VHF/UHF Power Transistor Amplifier Design", ITT Application Note, 1967, pp.15-16.

a high-beta transistor;[8]

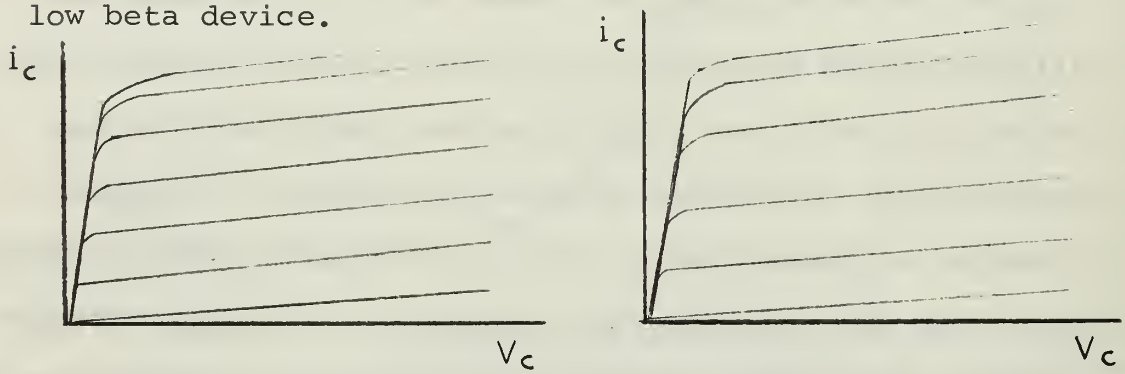
a. Less tendency to oscillate at a lower frequency.

b. Less difficult to maintain a constant bandwidth as a funtion of circuit layout.

c. Higher maximum power output (but lower power gain).

d. More dc bias stability.

Figure 12 shows a low-beta transistor and a high-beta transistor, to demonstrate the increased linearity of the low beta device.



Low linear beta high beta

Figure 12. - Linear and non-linear characteristic curves.

In audio-frequency power amplifiers; push-pull circuits are useful to obtain high power outputs with little distortion. The extreme difficulty in obtaining well-balanced, out-of-phase driving signals (of high power) precludes push-pull operation from high-frequency power amplifier consideration.[9]

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8 J. G. Tatum; "Microwave Transistor", ITT Application Note, 1967, p 14.

9 R. C. Hejhall; "Getting Transistors into Single-Side-band Amplifiers", Motorola Semiconductor Products, Inc., Application Note AN-150, July 1967, p 5.

The most common transistor configuration, in high-frequency, high-power, transistor, linear-power amplifiers is common-emitter. This is true for both class A driver stages and class B power stages. In the case of class A driver stages, the common-emitter circuit's phase reversal is desirable for negative feedback broadbanding.[10] In the case of class B power stages, a slightly higher power gain can be obtained with a common-base configuration, but there are considerable problems with high-frequency stability in common-base power amplifiers.[11]

For normal communication purposes it is desirable to have the intermodulation distortion at minus thirty decibels (-30db) or lower. The intermodulation distortion level desired, in conjuction with the bandwidth desired and the operating frequency range will determine the Q of the output circuitry. The output circuitry must fulfill the following functions:

a. Match the output impedance of the transistor to the load (for discussion purposes let the load be a commonly used value, 50 ohms). Normally, for high-power transistors, the output impedance is less than 50 ohms and is ca-

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10 P. Kolk; "Design of Wideband Transistor Amplifiers", Radio Corporation of America, Application Note SMA-7, 1961, p 3.

11 R. Minton; "Design of Large-Signal VHF Transistor Power Amplifiers", Radio Corporation of America, Application Note SMA-36, July 1964, p 1.

pacitive.

b. Reject harmonic frequencies. (Note that the frequencies to be rejected are all higher than the desired frequency; hence the output circuit will be basically a low-pass circuit.)

c. While rejecting harmonics, the circuit must either be broadband enough to allow carrier frequency changes or be tunable. Normally class B output stages are tunable. In class A stages, where b. above is not applicable, the circuit is normally broadband, with the "best" match at the high-frequency limit of the desired bandwidth. The best match is made at the high-frequency end, to counter the fall off of gain at higher frequencies that is normally found with transistors.

If the desired intermodulation distortion figure cannot be obtained with a single tuned stage, two or more tuned stages can be employed. However, this will cause more tracking problems in tuning.

2. Preliminary Design Discussion.

There are essentially three design approaches for transistor amplifier design. They are listed in the order of the discussion to follow. The three are discussed as applicable to high-frequency, high-power, transistor linear-power amplifiers only.

a. Design from equivalent circuit characterization.

b. Design from graphical techniques.

c. Design from two-port parameters.

Design from the equivalent circuit characterization of the transistor is usually the most difficult method, though it does have certain advantages.[12] Rarely, if ever, is sufficient data presented by the manufacturer to establish all values for an accurate equivalent circuit, and measurements are very difficult. The hybrid-pi equivalent circuit is the most useful equivalent circuit for common-emitter high-frequency transistors.[13] Being aware of the physical model of the transistor does aid one in understanding the behavior of the transistor.

The graphical technique of plotting a load line on the characteristic curve (collector current versus collector-emitter voltage for various base currents) of the transistor

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12 J. G. Linvill and J. P. Gibbons; Transistors and Active Circuits, McGraw-Hill, 1961, p 219-221.

13 R. Minton; "Design of Large-Signal VHF Transistor Power Amplifiers", RCA Application Note SMA-36, p 1.

has applicability for high-frequency, high-power transistors, though obtaining such a curve would be quite difficult.[14] The curve need not be obtained, however, to enable one to use the concept as an aid in understanding the high-frequency, high-power transistor. The purpose of this discussion is for conceptual understanding and is not an argument for high-frequency curve tracing.

Figure 13 shows the low-frequency and high-frequency characteristic curves. The curve to the right of the dotted line (safe operating area limit) could not normally be obtained without running a high risk of destroying the transistor.

The tendency of th high-frequency curves to be lower in magnitude and moved slightly to the right is frequency dependent and is due to the effect of an increased voltage drop across the base sheet resistance. ($r_{bb'}$ in hybrid-pi equivalent circuit, figure 5)

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14 It is not recommended that a characteristic curve be obtained for high-frequency, high-power transistors without carefull attention to the safe operating area published with the transistor data sheet, and adequate heat sinking.

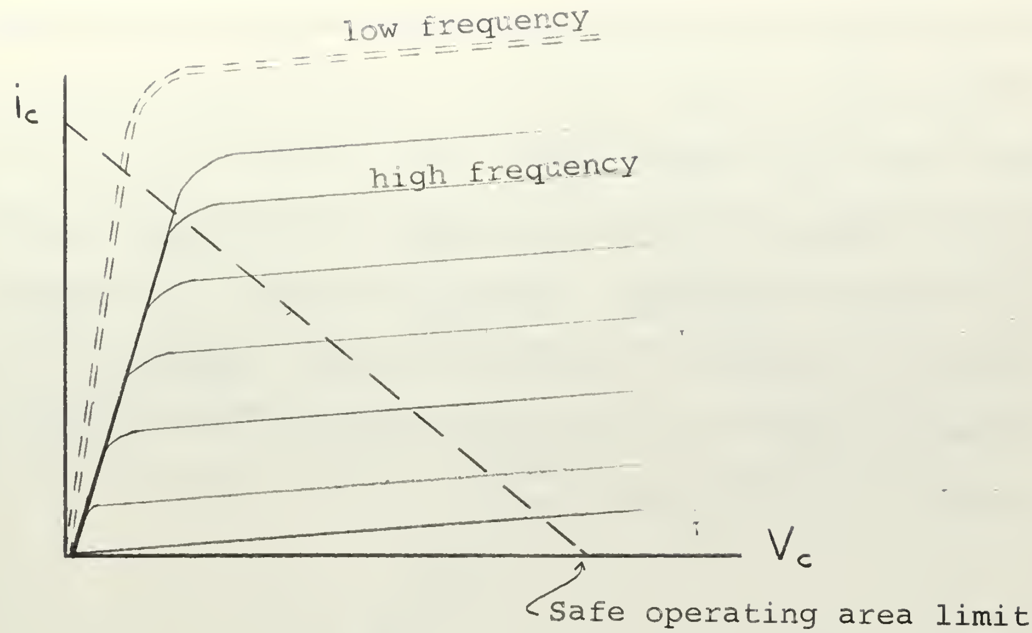

Figure 13. Low frequency and high frequency characteristic curves.

The increased voltage is due to the same driving current being required plus an increased current through the shunting capacitance ($C_{b'e}$ in the hybrid-pi). If a maximum power output versus frequency curve is contained in the data sheet, it could be expected to correspond roughly to the slumping of the high-frequency characteristic curve.

A load line is not drawn in the conventional manner, as the output impedance is frequency dependent, being of large magnitude (and reactive) except in the pass-band region. Load lines are needed for different purposes. The collector current, as shown in figure 14, justifies conventional load line, $R_{L'}$. The resonant phenomen, producing an output voltage as shown, justifies a load line R_L, where

the slope of load line R_L is one half that of load line $R_{L'}$, or

$$R_L = 2R_{L'} .$$

$R_{L'}$, of course, is the resistance of the output circuit at the carrier frequency. Normally, however, the high-frequency, high-power transistor's output impedance is complex, so that for maximum power transfer the output will need to be a conjugate match.

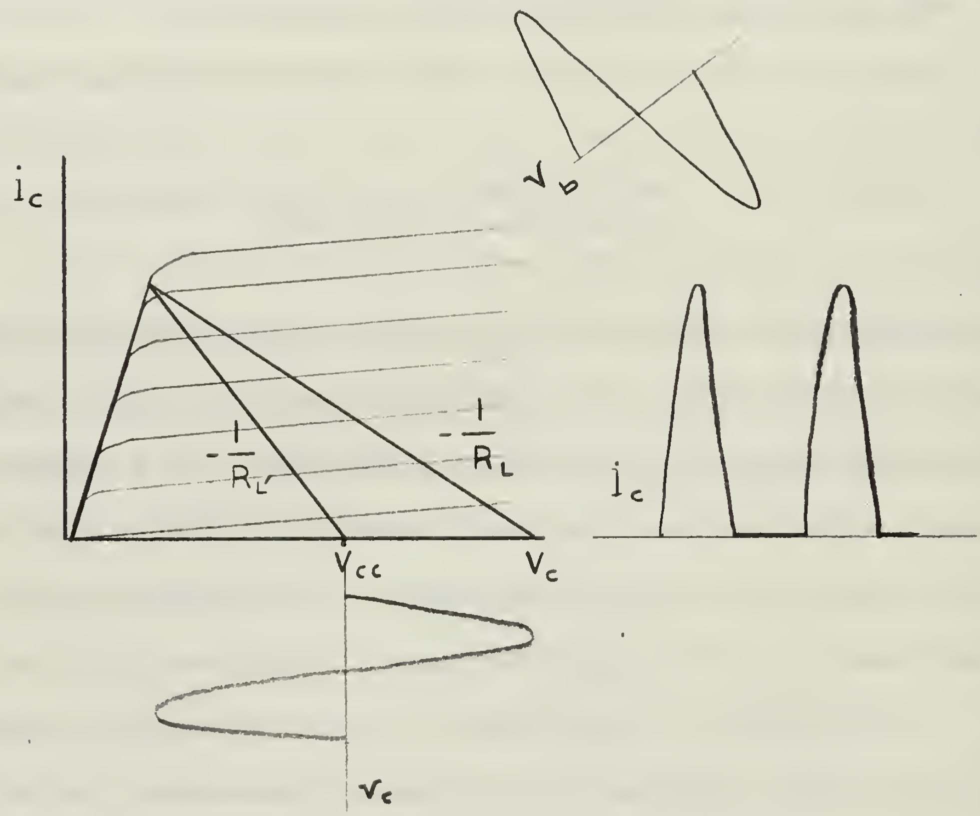

Figure 14. - Voltage and current waveforms for a Class B biased transistor.

Two-port design techniques (such as h, y or z parameters), as used in small-signal transistor design cannot be used if for no other reason, then simply because the large-signal two-port parameters cannot be determined. There have been attempts to modify small-signal parameters for use as design starting points for large-signal amplifiers. Such modified parameters and rules may be suitable for very limited purposes, but are inadequate for general use.

3. Measuring Parameters.

There are at least five widely known and commonly used pieces of equipment for measuring transistor y-parameters.[15] It is also fairly easy to construct elementary jigs to make y-parameter measurements on many RF bridges. Such y-parameter measuring equipments are indispensable for small-signal designing, but of very limited use in large-signal designing.

Motorola's method of determining large-signal impedance data is to operate the device in a circuit, peak the circuit's adjustable elements (for maximum power out), then remove the transistor without disturbing the circuit's impedances and measure the circuit. The claim is made that since maximum output power was being obtained, for a given input power, the circuit must have been the conjugate match to the transistor.[16] Large-signal impedance data published by Motorola and ITT are for class C biasing, though the data seems to be almost directly applicable to class B circuits. Output resistance usually is not given, but is left to be computed by the designer from the equation,

$$R_{out} = \frac{(V_{cc} - V_{sat})^2}{2\, P_{out}}$$

This equation is, of course, correct for class B circuits,

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15 O.A. Kolody and R.R. Langendorfer, "Measure Transistor y-Parameters", Electronic Design, 30 Aug. 1966, pp.54-59.

16 R. C. Hejhall, "Systemizing RF Power Amplifier Design", Motorola Application Note 282, 1967.

as is $C_{out} = 2C_{ob}$. (These equations are discussed in the following part.)

The published large-signal data is a useful aid in designing a class B power amplifier. It is unreasonable to expect one to design a circuit, and then measure the large-signal parameters in the Motorola fashion, since the impedance information will have then lost its timeliness.

For devices that do not have large-signal, class C impedance characteristics readily available the following (parts 4 and 5) is proposed.

4. The Output Impedance.

The value of R_L can be determined from the voltage swing available and the power out for the transistor in that particular situation. The voltage swing available is the collector voltage less the saturation voltage.[17]

$$R_L = R_{out} = \frac{(V_{cc} - V_{sat})^2}{2\ P_{out}}$$

Where R_{out} indicates the output, parallel resistance, and shall be used during the remainder of this thesis. The resonance phenomenon accounts for the voltage swinging as far above V_{cc} as it does below it. (Notice the manifestation of the rule that V_{cc} must be less than one half V_{CEO}.)

This large voltage swing has an effect on the voltage-sensitive collector capacitance.[18]

$$C_c = C_{ob} = \frac{1}{(V_{ce})^{\frac{1}{2}}} \left(\frac{\epsilon \epsilon_o q\ N_c}{8} \right)^{\frac{1}{2}}$$

Minton, in reference 19, computes the average value of the output capacitance during large voltage swings to be twice C_{ob} or,

$$C_{out} = 2\ C_{ob}\ .$$

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17 This equation is widely published. Proper credit for its origin is unknown.

18 H. Wolf; "Recent Advances in High Frequency Power Transistors", Solid State Design, June 1963, pp 21-28.

19 R. Minton; "Design of Large-Signal VHF Transistor Power Amplifiers", Radio Corporation of America Application Note SMA-36, July 1964, pp 14-15.

(This is the output capacitance, in parallel with R_{out}.)

C_{ob} is generally provided in specification sheets of high-frequency, high-power transistors. Thus it is possible to completely design a matching output network for a specified supply voltage and power output. V_{sat} can be assumed to be 1 to 1.5 volts without introducing sizable error, or can be ignored completely with 28-volt or higher supply voltages.

5. The Input Impedance.

By lumping together the base-emitter capacitance and the base-collector capacitance, as modified by the Miller effect, the input circuit can be approximated as shown below. $r_{bb'}$ is a constant for a given transistor and is

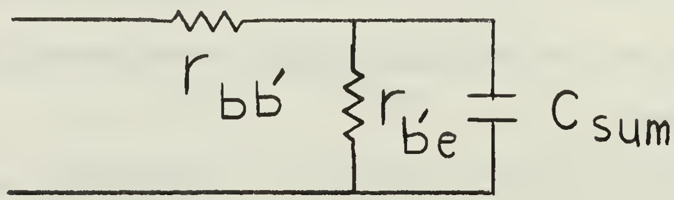

Figure 15. - Approximate input circuit for a high-frequency, high-power transistor.

measurable, if not provided in the transistor's specification sheet. $r_{bb'}$ may be measured on any small-signal transistor measurement device that has sufficient frequency range, a proper mounting device (common-emitter), and the correct impedance range. The measurement is made at frequencies somewhat above f_T . Not shown in figure 15 is the ever present lead inductance. Measurement of $r_{bb'}$ is made at the frequency at which the lead inductance is in series resonance with C_{sum}.[20] As the data in the appendix tends to support, an approximation of $r_{bb'}$ may be obtained even without such resonance.

At normal operating frequencies for the device, the input impedance may be considered to be $r_{bb'}$ in series with

20 R. P. Abraham and R. J. Kirkpatrick; "Transistor Characterizzation at VHF", Proceedings of the National Electronics Conference, 13, 1957, pp 385-402.

the sum capacitance.[21]

Also from reference 22,

$$X_{C_{b'e}} = r_{e'} \quad @ \quad f_T$$

and,

$$r_{b'e} = \frac{1}{1 - \alpha_o} r_{e'} = \frac{1}{1 - \alpha_o} \frac{k\,T}{q\,I_E}$$

Since for a useful device α_o will be .9 or greater, $r_{b'e}$ will be much larger than $X_{C_{b'e}}$; hence $r_{b'e}$ may be ignored for input impedance calculations at frequencies above 0.1 f_T.

Conversely it may be concluded that at a lower frequency, the input impedance will be higher. Operation at frequencies very low compared to f_T is not recommended, due to the possibility of second breakdown.

I_E can be determined for ideal class B operation by,

$$I_E = \frac{2}{\pi} \frac{P_o}{V_{cc} - V_{sat}},$$

so $r_{b'e}$ may be determined.

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21 The author has been unable to establish any wieldy rule to determine the sum input capacitance.

22 L. P. Hunter, Handbook of Semiconductor Electronics, McGraw-Hill, 1962, pp 12-10 to 12-13.

6. Biasing for Class B.

There are several design aspects that may be considered almost independently, and each, independently, may be considered of critical importance. After correctly designing with respect to these aspects, they need not be of further concern. If they are improperly designed, however, device failure may result, probably while the designer's attention is on other aspects of the problem. Heat sinking and biasing are such design aspects.

The requirements for the base biasing network are:

a. To provide a voltage equal to the threshold voltage of the base-emitter junction, regardless of the transistor's temperature, which will vary.

b. To provide sufficient power (current) to maintain the above voltage for all drive conditions. The input signal will be rectified across the base-emitter junction, tending to lower the base voltage.

As discussed in section II, a diode placed on the high-frequency, high-power transistor chip and connected to an external unity-voltage, current amplifier is a good solution to the problem.

In most situations a bypassed emitter resistor in conjunction with a RF-choked, forward-biased diode will be required to achieve the desired stability. The emitter resistance should be kept as small as possible to avoid loss of efficiency. Large bypass capacitors will be required. The reactance of the RF choke should be ten-to-twenty times

the input impedance. In some cases, because the emitter is connected directly to the heat sink, it may be desirable to have the lower end of the emitter resistor below ground potential. A practical rule of thumb might be to raise the bias voltage from class C toward class B until the desired intermodulation distortion value is obtained.

7. Combining Transistors.

The purposes for combining high-frequency, high-power transistors are to increase output power and/or to improve reliability. A two-fold increase in power output cannot be obtained by paralleling two transistors directly, as they would not share the load equally. This is precisely the reason why the emitter-ballasting resistance was added to the multi-emitter devices. Proper paralleling schemes make use of broadband transformers[22] or other combiners.[23]

Increased reliability is obtained by requiring less power output from each transistor. While being driven with less power and providing less power output, the transistors are capable of withstanding more adverse conditions, such as high ambient temperature or load mismatch.

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22 E. Arvonio; "Use of Broadband Toroid Transformers in High-Frequency Linear Amplifier Applications", US Naval Air Development Center, Johnsville, Pennsylvania, NADC-EL-6356, October 1963.

23 C. H. Wood and A. W. Morse; "Solid State HF Linear Amplifier", Westinghouse Electric Corporation, RADC-TR-66-579, (U.S. Government Nr. AD 822 313), October 1967, p 2-8.

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APPENDIX.

1. $r_{bb'}$ Measurement.

$r_{bb'}$ measurement was accomplished on a GR 1607-A transfer-function and immittance bridge. The local oscillator trap was not used below 300 Mhz due to excessive attenuation (40 db at 500mhz plus 6db per octave below 500 mhz).

40340 Transistor; (These measurements were taken with the plastic cap removed from the transistor common-emitter mount, and no bias voltages.)

Table 2. 40340 $r_{bb'}$ data.

Mhz	Impedance (ohms)
150	$<0.5 + j0.0$
175	$0.5 + j1.0$
200	$0.5 + j2.5$
225	$0.5 + j3.0$
250	$0.5 + j3.5$
300	$0.5 + j4.5$

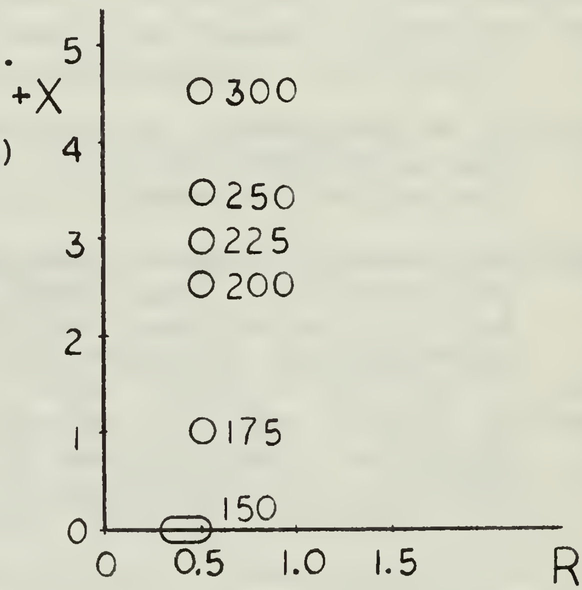

Figure 16. 40340 $r_{bb'}$ data.

$r_{bb'}$ determined to be less than 0.5 ohms.

(f_T was not specified in the data sheet, but is estimated to be 150 mhz.)

40341 Transistor; (These measurements were taken with the plastic cap removed from the transistor common-emitter mount, and no bias voltages.)

Table 3. 40341 $r_{bb'}$ data.

Mhz	Impedance (ohms)
150	$<0.5 + j0.5$
175	$<0.5 + j2.5$
200	$0.5 + j3.0$
225	$0.5 + j4.0$
250	$0.5 + j4.5$
300	$0.5 + j5.0$

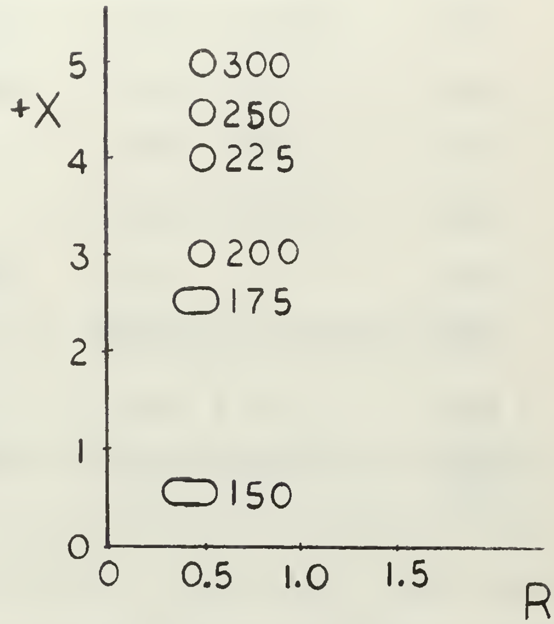

Figure 17. 40341 $r_{bb'}$ data.

$r_{bb'}$ determined to be less than 0.5 ohms.

(f_T was not specified in the data sheet, but is estimated to be 150 mhz.)

2N2876 Transistor; (These measurements were taken with a mounting block that contained 8mm leads connecting the transistor to the common-emitter mount.) $V_c = 10v$, $I_c = 5ma$. $r_{bb'}$ is estimated to be 4.0 ohms. (f_T is specified as 200 mhz in the data sheet.)

Table 4. 2N2876 $r_{bb'}$ data.

Mhz	Impedance (ohms)
300	14.0 + j29.5
400	7.5 + j 9.0
500	4.0 + j19.5
600	4.0 + j24.0
700	4.5 + j36.0
800	4.5 + j43.5
900	7.0 + j59.0
1000	7.0 + j65.0

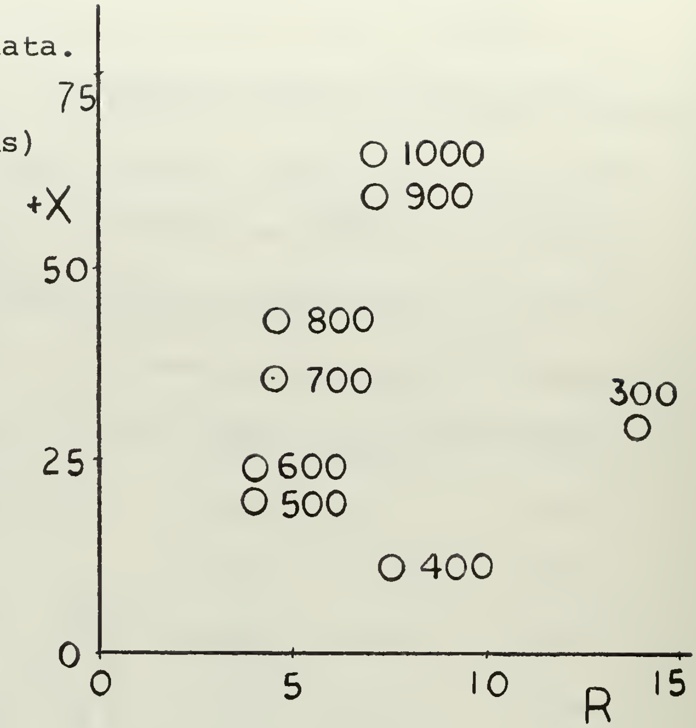

Figure 18. 2N2876 $r_{bb'}$ data.

The transistor mounting scheme introduced excessive inductance, making it impossible to resonate C_{sum} with the lead inductance at frequencies above f_T. Correlating similar measurements made with 40340 and 40341 transistors indicates the minimum real part of the impedance is nearly $r_{bb'}$. Subsequent destruction of the 2N2876 transistor prevented re-measurement without the mounting block.

40340 and 40341 transistor data with the mounting block is given below, for comparison with 2N2876 transistor data.

Table 5. Data for comparison to 2N2876.

Mhz	40340 (ohms)	40341 (ohms)
150	3.0 + j 5.0	2.5 + j 7.0
200	1.0 + j 7.0	0.5 + j 7.5
300	1.5 + j10.0	1.0 + j 8.5
400	2.0 + j20.5	1.5 + j19.0
500	2.0 + j29.0	1.0 + j25.5

2. Large-Signal Transistor Parameter (Impedance) Measurement.

The technique of measuring the large-signal transistor parameters was to operate the transistor being tested in a circuit under the desired test conditions, and then to tune the input and output matching networks for peak power out. With peak power out, for a given power in, a conjugate impedance match is obtained at the input and output of the transistor. Then the dc power supplies are replaced with short circuits, the input feed line (to which the power amplifier-driver is matched) is replaced with its characteristic impedance, and the transistor is removed. None of the above steps should alter the conjugate impedance match. The impedance can then determined with a RF bridge. (HP-1606 and Wayne-Kerr B-601 were used to determine input and output impedances respectively.)

Practical difficulties encountered were:

a. Larger variable capacitors were needed. (170pf

was the largest available, of suitable physical size).

b. The shielding of the transistor circuit made it impossible to replace elements quickly.

c. The power amplifier-driver was a tuned vacuum-tube stage, and had to be pi-matched to the input impedance for each power setting.

d. Pure resistive loads capable of dissipating the large quantities of power could not be obtained.

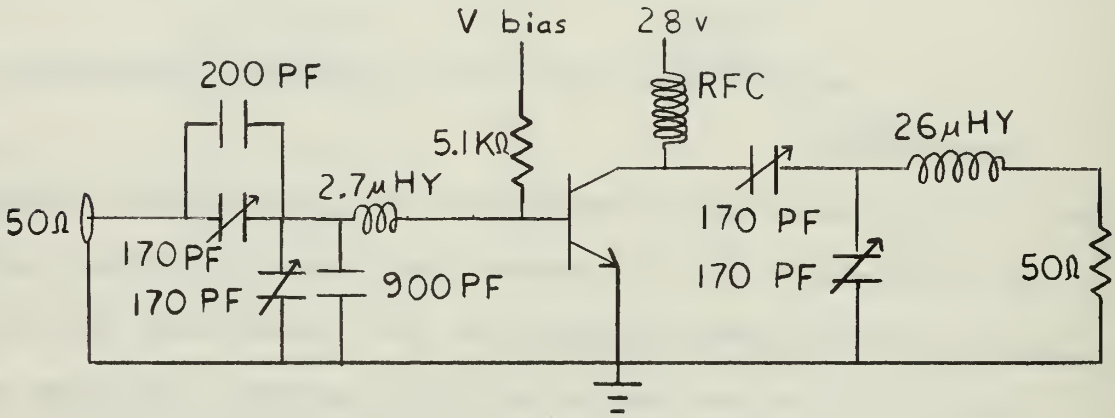

Figure 19. 2N2876 class A circuit, 3mhz.

Table 6. Large-signal transistor measurements, 2N2876, class A, 3mhz.

I_c	In (series)		Out (parallel)	
ma	ohms	pf	ohms	pf
100	5.5	5,000	590	46
200	5.0	10,000	350	36
300	5.5	10,000	260	54

2N2876 Class B Circuit, 3Mhz.

The above circuit was simply biased class B. Preliminary adjustments were made to match impedance, but the transistor was destroyed (by removal of input signal without first removing the base bias voltage) prior to final adjustments. A rough determination indicated that the transistor had nearly the same input impedance, but a lower output impedance under class B conditions.

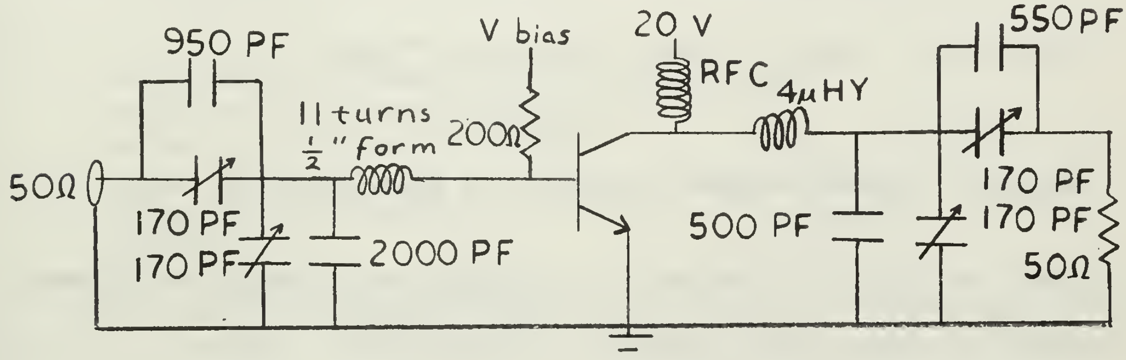

Figure 20. 40340 Class B Circuit, 3Mhz.

Table 7 Large signal transistor measurements, 40340 class B.

Power in (ac); 0.46 watt Power out (ac); 15.7 watt

Power in (dc); 30.0 watt Efficiency; 52%

Impedance;	In (series)		Out (parallel)	
	ohms	pf	ohms	pf
	5.8	4,500	22	1,700

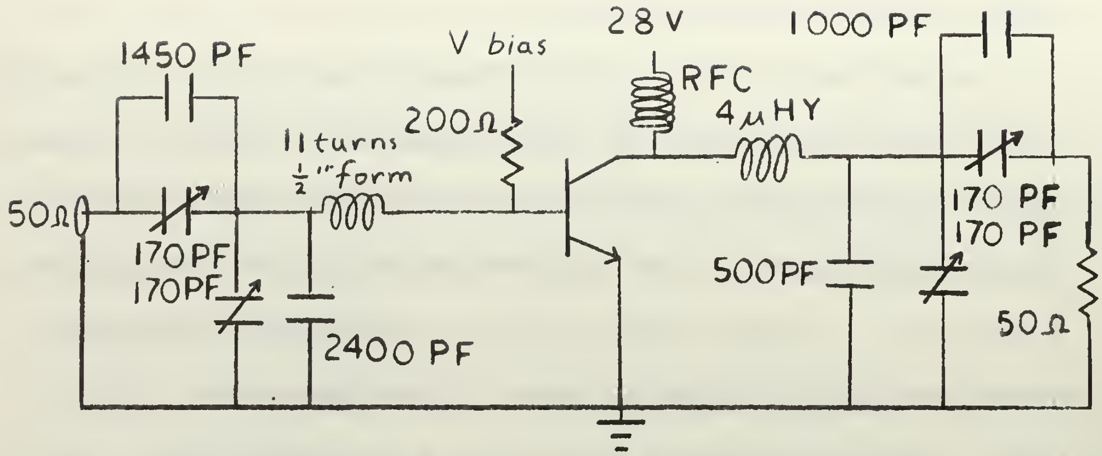

Figure 21. 40341 Class B Circuit, 3Mhz.

Table 8. Large signal transistor measurements 40341 Class B.

Power in (ac); 0.96 watt   Power out (ac); 26.0 watt

Power in (dc); 40.0 watt   Efficiency; 65%

Impedance;	In (series)		Out (parallel)	
	ohms	pf	ohms	pf
	5.3	3,400	34	900

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13. ABSTRACT

This thesis discusses in detail the physical and electrical characteristics of high-frequency, high-power transistors, and why class B amplifiers are necessary for linear power amplification of signals containing more than one frequency.

Linear power amplifier design is contingent upon having a suitable design technique. "Suitable" often means being able to determine parameter values called for by that technique. Conjugate impedance matching is a suitable technique and three of the four parameter values can be accurately determined. In some cases manufacturers provide data for this technique., titled "Large-signal parameters".

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14 KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
High-Frequency, High-Power Transistor Linear Power Amplifier.						
Radio-Frequency Power Amplifier						
Radio-Frequency Linear Power Amplifier						
Transistor r-f Linear Power Amplifier						
Calss-B Linear Power Amplifier						
Transistor Power Amplifier						
Transistor Large-Signal Amplifier						